

DIRECT-CONVERSION CW TRANSCEIVERS

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THE CONTENTION of many that "we live in a complex age" is no better illustrated than by merely glancing at the circuit diagram of a Yaesu FT757, or similar, hf transceiver. Quite understandably, the majority of radio amateurs have long ago buried any dreams of operating an entirely home-constructed station, and the "do it yourself" aspect of our hobby is confined almost entirely to the building of ancillary equipment and the erection of antennas. The problem—assuming that the present situation is, at least, to a degree undesirable—clearly deserves analysis. Leaving aside for a moment the ramifications of contemporary technology such as microprocessor control and digital frequency synthesis, I believe that it was the emergence of the transceiver itself which put paid to much home-construction. There was, lest we forget, a time not so far distant when transmitters and receivers existed only as separate entities. Although relatively few amateurs have ever constructed a double- or triple-conversion superheterodyne receiver, a significant number of operators would previously have built their own hf transmitter for use in conjunction with a commercial receiver. The dawn of the transceiver, bringing the inestimable advantage of single-knob tuning, plus many other unified, even automatic, control functions, changed things completely and irrevocably. Many potential constructors have since found that the necessity to build everything into one small box can represent a psychological hurdle which proves impossible to assail.

Finally, the adoption of ssb as the prime mode for hf telephony aggravated matters further, by throwing a whole generation of amateurs who had been weaned on no-nonsense a.m., into a state of confusion and bewilderment. Optimists may take heart, however, from the realization that the complexities of ssb generation can be completely ignored if the constructor sets his sights on the building of a cw-only transceiver. Not only will such a transceiver be considerably simpler to construct, but it will also provide an extremely high degree of communication effectiveness, for it is still an acknowledged fact that cw is the superior mode for weak-signal and dx working.

The remaining questions concern the application of sophisticated digital-frequency synthesis techniques and microprocessor control. Quite simply, these most recent developments have not brought about any fundamental change in the operating principles of contemporary hf transceivers. This is because the signal chain, comprising various amplifiers, mixers and filters, remains totally analogue and we have not yet arrived at a point where the digitization of the signal itself, and the consequent application of digital filtering etc, becomes entirely practicable. Obviously, the partial digitization of the vfo "side chain" has produced many operational advantages—digital readout and spot frequency memories, for instance—but as these convenience features are largely peripheral to the core of the transceiver function, they may legitimately be ignored by constructors striving towards an economic and uncluttered design.

Many of us learned about trf (tuned radio frequency) receivers while studying for the RAE, and it is worth remembering that in the 'twenties and early 'thirties the trf held sway over the superheterodyne. More recently, a technique allied to the trf principle, direct-conversion, has aroused considerable interest among radio enthusiasts, particularly those searching for simpler, but nonetheless high performance designs. A great advantage of the d-c receiver is that it may be converted painlessly into a practical cw transceiver featuring single-knob tuning and automatic transmit/receive changeover.

Direct-conversion

Fig 1 shows the essential elements of a direct-conversion receiver, and most readers will be familiar with the principles involved. Briefly, the local oscillator is tuned to a frequency very slightly offset from that of the wanted cw signal so as to generate an audible beat note whenever the transmissions

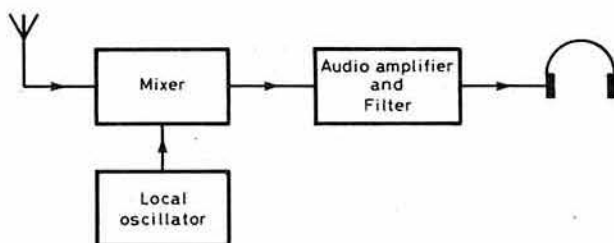


Fig 1. A direct-conversion receiver

carrier is present. By down-converting the signal from transmission frequency to audio in a single step, we produce a very elegant cw receiver which utilizes only one mixer and substitutes the normally expensive crystal or ceramic i.f. filter found in superhets with an audio filter that may comprise just a few CR networks and some cheap ic op-amps.

Inevitably, there is a penalty attached to such simplicity—the frequently-discussed "audio image". For any local oscillator setting, audible products will be generated by signals having carrier frequencies either slightly above, or slightly below the oscillator's frequency. This double response means that the bandwidth of the d-c receiver is effectively twice that of an equivalent superhet. Also, as the receiver is tuned across the band each transmission is heard twice—once when the vfo frequency is lower than the signal, and again as the oscillator is incremented to a point just above the signal.

The audio image is clearly a bugbear of direct-conversion that must be taken into account by anyone contemplating the design and/or construction of such equipment. However, readers may rest assured that the d-c technique is entirely practicable, and the slight degradation in receiver performance should rightly be set against the simplicity and low cost achievable with direct-conversion.

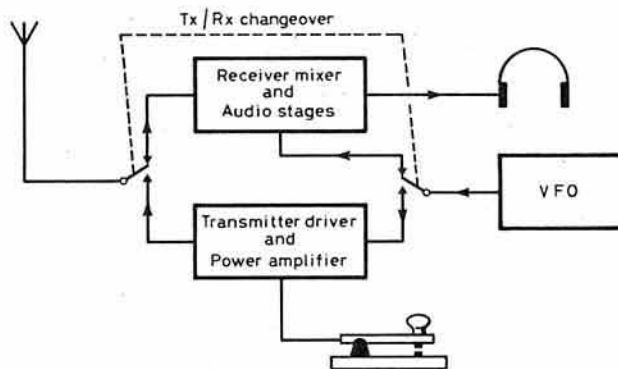


Fig 2. A direct-conversion cw transceiver in simple form

Fig 2 illustrates a basic cw transceiver featuring direct-conversion in the receiver. Because the vfo operates at signal frequency, much of the complexity associated with superhet rigs is avoided and transmit/receive changeover becomes particularly straightforward.

The remainder of this article is devoted to the description of practical circuitry developed for an hf direct-conversion cw transceiver. I have adopted a modular approach where each section, or functional block, within the rig is self-contained and may be built and tested separately. What follows will appeal more to the fairly-experienced constructor and experimenter. Precise layouts, component lists etc have been avoided. So has description of circuitry that may be taken for granted; eg the audio output stage and mains psu.

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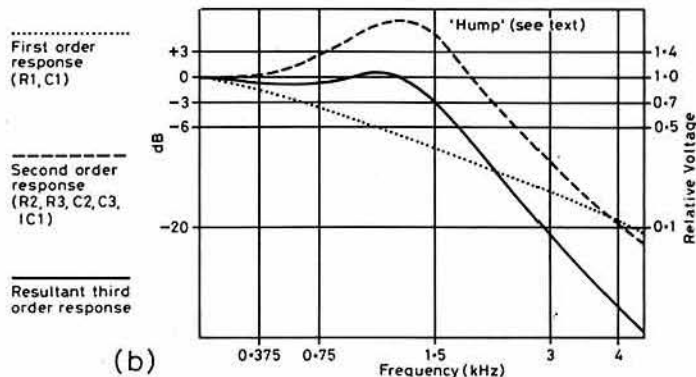
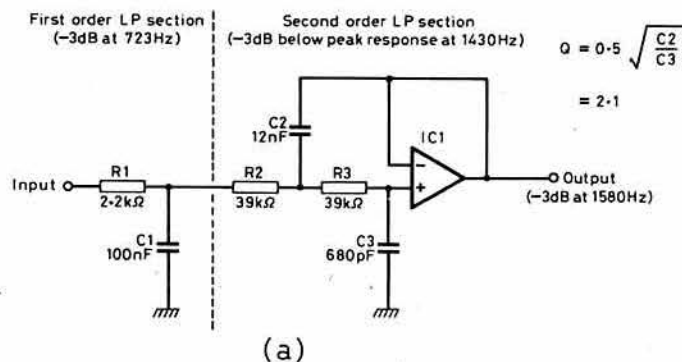


Fig 3. The third-order lowpass filter and associated response curves

Receiver

The direct-conversion receiver has perhaps suffered through having the label "simple" attached to it. Indeed, many designs featuring an incredibly low parts count have found their way into the amateur radio literature, and it is true to say that a complete d-c transceiver may be built using the veritable "handful" of junkbox components. Such ingenuity and thrifty enterprise is, of course, to be applauded, but it does have the unfortunate side effect of masking many fundamental design considerations. If, instead, we treat the direct-conversion receiver as being merely a straightforward engineering solution to the problems of hf reception, considerable headway can be made. After all, the d-c receiver, although possessing fewer stages than the superhet, does still require a local oscillator, various filters, an amplifier and a mixer. Therefore, all the performance criteria associated with such functions; eg oscillator stability, filter bandwidth and shape factor, amplifier distortion and noise, mixer intercept figures etc, are entirely relevant to the d-c receiver also.

Let us consider the filter first. The mixers output will comprise a very wide spectrum of products stretching from virtually dc (direct current this time!) at the low end, to frequencies of the megahertz order at the high end. The human ear is only sensitive to frequencies within the range 30 to 20,000Hz, and so the subjective bandwidth of a d-c receiver, without any form of filtering, could be said to be approximately 20kHz. However, this bandwidth is far wider than that necessary for ssb telephony, which requires no more than 2.7kHz, let alone cw. Also it is most important to realize that a high-level mixer product at, for instance, 35kHz, although inaudible in itself, might cause overloading of an amplifier and, therefore, the generation of both second- and third-order products associated with such distortion.

In the typical junkbox rig referred to earlier, a single pole lowpass filter, interposed between the mixer and audio amplifier, often constitutes the only post mixer filter used within the entire receiver. Although better than nothing, such an elementary filter cannot possibly provide the shape factor desired in a high-performance receiver. Frequencies up to 10kHz will be plainly audible, a gross disadvantage in itself, and for reasons just discussed the dynamic range of the receiver will be considerably compromised.

Before investigating better and necessarily more complex filters, it is important to decide what constitutes the optimum response shape. The constructor's first instinct, in the case of a cw transceiver, might be to opt for a multi-stage bandpass filter resonant at around 800Hz and possessing a bandwidth of no more than 300Hz. While it is true that such a filter could provide excellent isolation of the wanted signal, do we really want to be limited to such a narrow bandwidth? The problem, clearly, is that with particularly sharp filtering, tuning of the receiver becomes rather difficult and the limited range of frequencies appearing at the receiver's output creates an acute sense of isolation from the rest of the band.

The second filter option involves the all-singing, all-dancing approach, where a tunable filter featuring variable bandwidth and variable centre

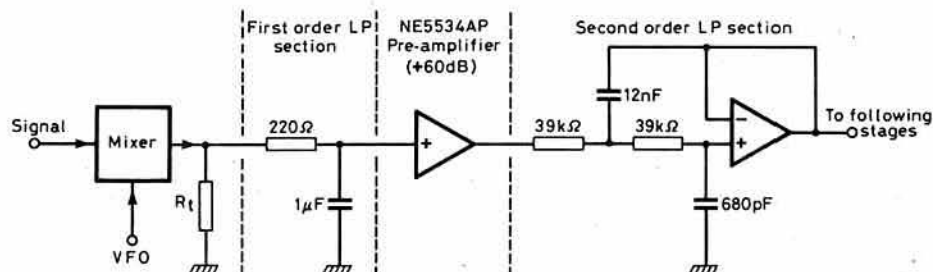
frequency is employed. This, I will concede, represents the ultimate. But there are many problems in implementing such a system. The circuitry required would be quite complex and any attempt to simplify the design is bound to result in compromised performance.

After much thought I decided on a multi-section, fixed-frequency active filter of the lowpass type having a -6dB response at approximately 1,400Hz. This filter, which provides over 60dB of attenuation in the first octave above cut-off, is supplemented by a simpler highpass filter which rolls off more gradually below 400Hz. The complete filter system therefore possesses a -6dB bandwidth of just over 1kHz and is capable of isolating cw signals very effectively, but without sounding too sharp. Needless to say, if a narrower bandwidth is ever desired, perhaps for dx chasing or contest work, then an add-on active filter of the type I have previously described [1, 2, 3] may also be employed.

Fig 3(a) shows the third-order lowpass section which is used as a building block for the main filter. R2, R3, C2, C3 and IC1 comprise a second-order lowpass filter that will work quite happily in isolation from R1 and C1. The Q or damping factor of the second-order section is determined by the ratio of C2/C3. A Q value of 1 would be obtained by simply reducing this ratio to 4:1, and the resultant amplitude response is then almost perfectly flat up to the cut-off frequency, at which point the filter begins to gradually roll off. However, if the Q is increased to around 2, the filter rolls off more sharply beyond the cut-off frequency, thus producing a better "close in" shape factor. The snag is that as the Q is increased beyond a value of 1, a resonance, or "hump" begins to appear in the amplitude response at a point just prior to cut-off. This irregularity could possibly be tolerated in a single-stage filter of moderate Q, but if a number of such filters are cascaded the hump grows proportionately in amplitude and we end up with something looking more like a rather sharp bandpass response. The problem is solved in the third-order filter of Fig 3 (a) by introducing a first order section comprising R1 and C1. The values of these two components are chosen so that the first-order section has a -3dB response approximately one octave lower than the second-order network. This has the most desirable effect of smoothing out the dreaded hump, and also increases the attenuation achieved beyond the cut-off frequency. Fig 3 (b) illustrates the technique by way of a graph, and reveals that an attenuation of around 20dB is obtained in the first octave above cut-off. In order to realize the desired performance, the complete lowpass filter employs three almost identical third-order stages cascaded.

Those readers who are versed in the mysterious and often highly-mathematical subject of filter design may wonder why a more developed, higher-order circuit has not been chosen. The reason for this lies in the fact that a seventh-order Chebyshev lowpass filter offering the same performance would demand the use of two per cent, or even one per cent tolerance components in order to guarantee a well-controlled response curve. Just as many components would probably be required, if only because the call for certain precise values might well dictate the use of series or parallel combinations of values actually available in the E12 or E24

Fig 4. The receiver module front-end in skeleton form



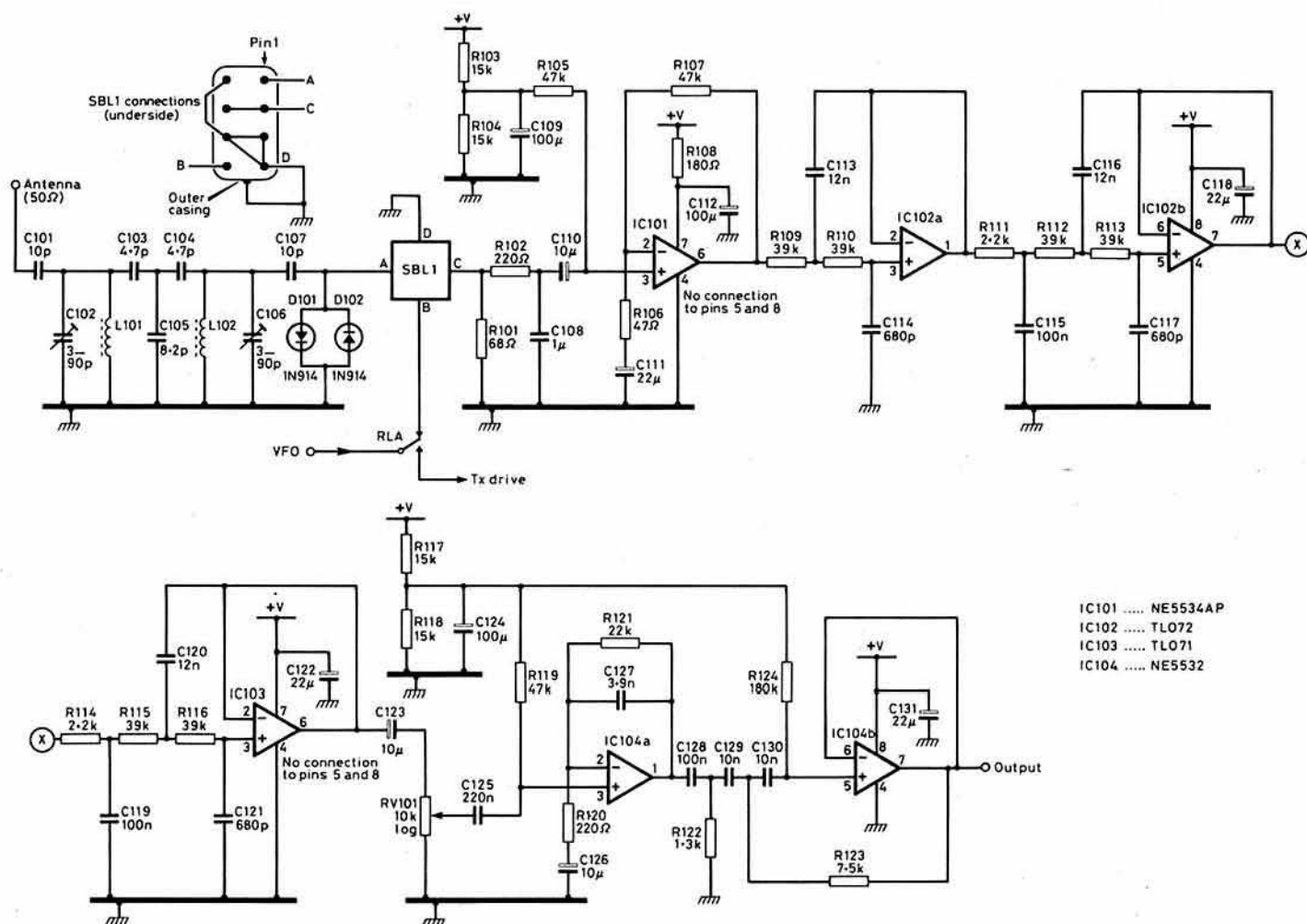


Fig 5. The receiver module for 14MHz, L101 and L102 are both 20t 22swg enamel wire, lightly wound on Amidon T68-6 dust-iron toroid cores

series. In contrast, the triple third-order filter works quite happily with standard five per cent E12 series resistors and capacitors.

As the active lowpass filter has a net gain of zero within its passband, considerable amplification must be provided by other stages in order to raise the microvolt level of weak signals so that they become audible in headphones or through a loudspeaker. Most of this gain is obtained by utilizing a high-performance bipolar op-amp, type NE5534AP. The suffix "AP" denotes selection for low noise, and in practice this means a noise voltage some 15dB lower than that generated by bi-fet op-amps such as the TL071.

Fig 4 shows the receiver front-end in skeleton form. The first "gain block", featuring an NE5534AP preamplifier, is interposed between the first-order and second-order sections of the initial third-order lowpass filter. The values of the first-order section are scaled by a factor of 10 (giving 220Ω plus 1μF) thereby ensuring very low noise and providing a better match to the mixer output port.

The mixer is a packaged Schottky diode ring type SBL1, and this device works best when all three ports (input, output and local oscillator) are terminated by a 50Ω resistive load. Resistor R1 is therefore incorporated so as to ensure that this condition is met in respect of the output port. Readers may question the use of a relatively expensive balanced mixer rather than one of the cheaper alternatives such as a simple dual-gate mosfet circuit. The central reason for this apparently extravagant choice is that single-ended mixers suffer from a deficiency best described as "envelope detection". Briefly, all mixers function by virtue of the fact that they are essentially non-linear. Unfortunately, the very non-linearity which enables the mixer to generate sum and difference products can also give rise to the partial rectification of strong carriers. In a superheterodyne receiver, the fact that a strong a.m. broadcast signal may be envelope-detected by the first mixer causes no real problems because the resultant audio is completely rejected by the i.f. filter(s) that follow the mixer. In a direct-conversion receiver, however, the envelope detected signal cannot be rejected for the simple reason that the receiver's filter is designed to pass audio frequencies.

The SBL1 mixer employed in my receiver module operates in the switching mode and employs "fast" Schottky hot-carrier diodes to reduce the transition time between the ON and OFF states. This performance feature, coupled with the doubly-balanced nature of the transformer coupled diode ring, provides a very high immunity to envelope detection.

The preamplifier voltage gain is 60dB; this may seem rather high, particularly when bearing in mind that there is only a single-pole filter section between the mixer output port and the amplifier's input. However, we must set against this gain two significant losses. First, there will be a two-section bandpass filter connected between the antenna and the mixer input port. This filter introduces a loss of approximately 6dB within its passband. Second, the SBL1 mixer has a conversion loss of around 6dB also. Subtracting these two losses from the preamplifier gain figure results in a net front-end voltage gain of 48dB. Also, it should be noted that the first-order filter section has a -3dB response point at roughly 700Hz. Therefore, the gain at this frequency drops to a quite manageable 45dB, and the receiver should therefore be capable of handling signals that develop 20mV rms across its 50Ω antenna input.

Fig 5 shows the complete circuit diagram of the receiver module, and Fig 6 supplements this by illustrating the system in block form, complete with gain distribution figures. Components C101 to C107 plus L101 and L102 comprise the signal frequency bandpass filter. The top coupling at each end of the filter provided by C101 and C107 has been calculated to present a 50Ω match. The filter is set-up by merely adjusting C102 and C106 (both miniature plastic foil trimmers) for maximum signal at the centre of the cw segment of the band. Although utilizing a fair number of fixed capacitors, this filter has the advantage of not requiring either tapped inductors or coupling windings.

The prototype receiver operates on 14MHz, but it is a simple matter to re-calculate the component values for other bands; all necessary design data being contained in Appendix 2 of *Solid state design for the radio amateur* [4], which I have employed for the development of the 14MHz filter. The two silicon diodes, D101 and D102, protect the mixer from excessive input

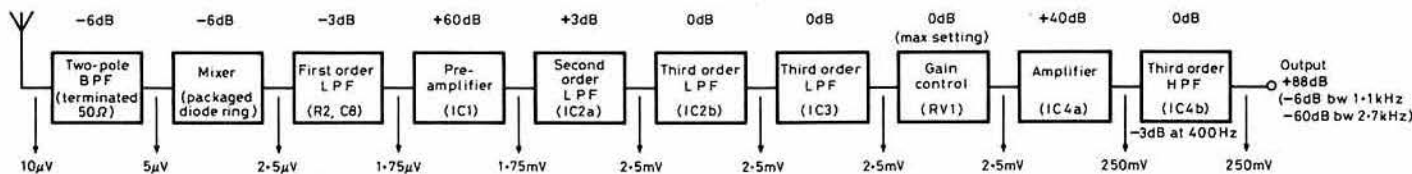


Fig 6. Block diagram of the receiver module showing gain distribution and voltage levels for a 10µV signal which produces a mixer output at 700Hz

voltage. VFO drive is routed to the mixer via a small changeover relay. On transmit, the relay coil is energized, thereby transferring oscillator drive to the transmitter module.

R101 terminates the output port of the SBL1 mixer, and R102 plus C108 form the first-order filter section discussed earlier. Series, rather than shunt feedback is employed around IC101, as this results in lower noise. The negative feedback is provided by R107 and R106. IC102a, R109, R110, C113 and C114 constitute the initial second-order section featured in Fig 4. The remaining third-order filters are built around IC102b and IC103. In order to avoid any confusion, it should be noted that IC102 is a dual op-amp (TLO72) whereas IC103 is a single op-amp (TLO71). RV101 is the gain control which precedes the second "gain block" (40dB) built around IC104a. IC104b, R122, R123, R124, C128, C129 and C130 form a third-order highpass filter with a -3dB response at 400Hz. This filter significantly attenuates the lower audio frequencies and thereby helps eliminate any residual mains hum and 100Hz ripple.

IC104 is a dual version of the NE5534 and was chosen for two reasons. First, it became apparent while testing the prototype receiver in conjunction with an LM380 audio output stage that hiss generated by IC104 was subjectively noticeable. Needless to say, such noise cannot be attenuated by reducing the volume setting, and so the lowest-noise op-amp is desirable in this position. Second, the NE5532 is fully capable of driving resistive loads down to 500Ω, whereas other op-amps can only manage, typically, 2,000Ω. This makes it possible to connect high impedance headphones directly to the output of IC104b. If headphones are to be used in this fashion, a 10µF blocking capacitor must be wired in series with the output.

Readers will no doubt have realized that the receiver does not employ any form of agc. Those who feel that agc is desirable, perhaps even essential, may rest assured that in practice the absence of agc has not proved particularly detrimental. Indeed, the fact that signals are presented to the listener with their true relative amplitudes preserved induces a satisfying sense of realism and makes a welcome change from the rather "processed" sound of most commercial receivers.

Many constructors will wish to add some form of audio power amplifier so that the receiver can be coupled to a loudspeaker. My prototype transceiver employs an LM380 ic output stage, but problems were experienced initially due to feedback caused by the LM380, which draws fairly high currents, modulating the receiver module supply rail. The solution is to feed power to the audio output stage via a separate voltage regulator. It is also important to incorporate the bias and supply rail decoupling electrolytics shown in Fig 5. My transceiver uses a 15V supply rail, and the psu, which is built into a separate screened case, incorporates two 7815 ic regulators. The main reason for choosing a voltage slightly higher than the virtually standard 13.8V was so that the transmitter output

stage (described later) could develop more power. However, all the circuits described should function perfectly well on 13.8V.

Fig 7 shows the circuit of a switchable 20dB attenuator for use between the output of IC104b and the audio power amplifier. This network complements the volume control (RV101) and enables the receiver's gain to be controlled with ease over a very wide range.

The fixed resistors employed in the receiver module are standard five per cent tolerance carbon or metal film types of 0.33W rating. Non-electrolytic capacitors used in the audio filter sections are five per cent tolerance

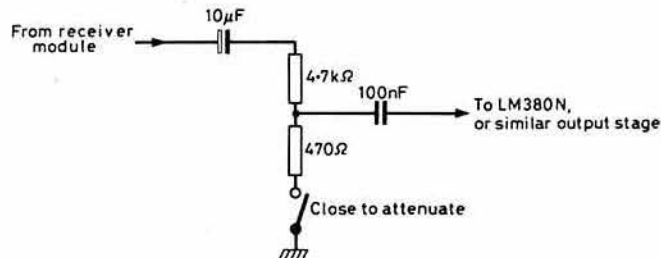


Fig 7. A switchable 20dB attenuator

polyester types (Siemens B32560 and B32561), except the 680pF value, which is polystyrene (Siemens B31310). Other types of capacitor can be employed (eg Mylar) so long as they are of five per cent tolerance. Ceramics are an exception, however, and should be avoided.

The receiver module possesses a sensitivity and signal-to-noise ratio which should prove adequate for all applications on bands from 1.8 to 7MHz. It may, however, be possible to improve the performance on the higher frequency bands by the addition of an rf amplifier. A single jfet amplifier stage, suitably integrated with the input bandpass filter network, might be tried. RF gain achieved in this way should be restricted to between 10 and 15dB. It will also be a good idea to incorporate a bypass switch for the amplifier, or alternatively, a switchable input attenuator.

VFO

Many direct-conversion transceiver designs published in recent years have relied on crystal control in order to achieve simplicity and guaranteed frequency stability. A limited tuning capability can be provided by "pulling" the crystal, and much ingenuity has been applied by the originators of vxo circuits in an attempt to bridge the gap between crystal and vfo control.

Few would argue, however, that a properly-designed vfo is the only

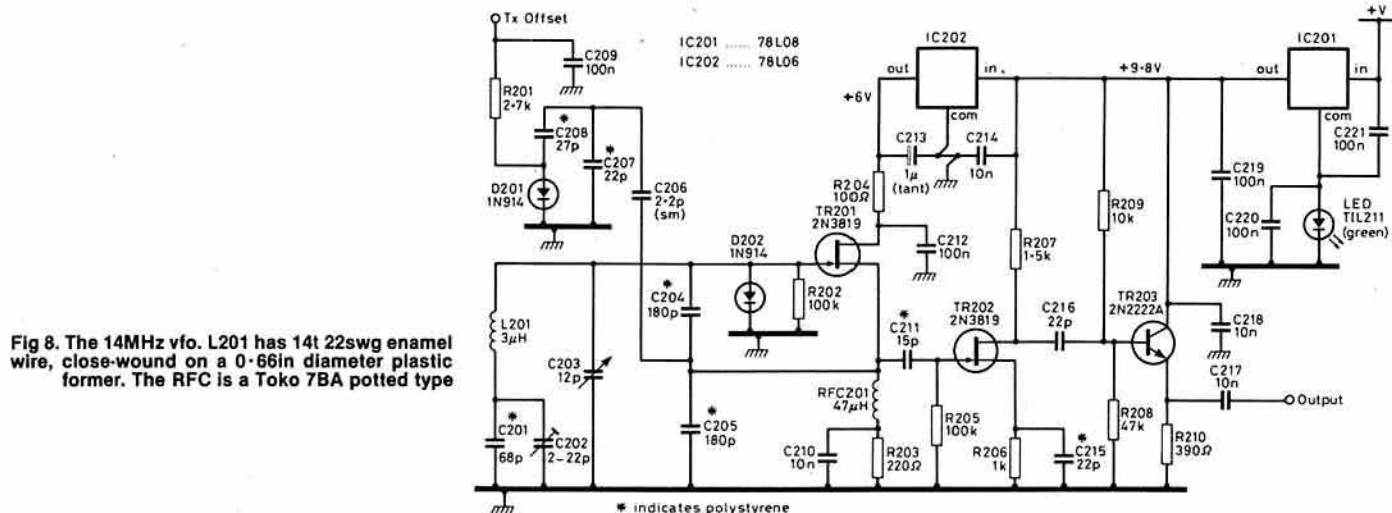


Fig 8. The 14MHz vfo. L201 has 14t 22swg enamel wire, close-wound on a 0.66in diameter plastic former. The RFC is a Toko 7BA potted type

serious alternative to frequency synthesis for a rig intended to cover the whole of a particular hf band, or even just the band's cw segment. Having opted for a vfo, I played safe by adopting the tried-and-tested Clapp arrangement, which is essentially a series-tuned Colpitts oscillator. Fig 8 gives the circuit, and there are few surprises here. IC202 generates a stable 6V supply for the oscillator (TR201) and IC201, working in conjunction with a green l.e.d., provides a 9.8V rail for the amplifier and output buffer (TR202, TR203). C203 provides tuning over a range of 250kHz and allows casual monitoring of ssb transmissions higher up the band (although it must be borne in mind that the audio filter employed in the receiver module is rather narrow for correct reproduction of speech).

C202, a plastic film trimmer, allows band-edge setting after final assembly of the vfo in a rigid, screened box. A slow-motion drive for C203 is, of course, essential, and a reduction ratio of about 30:1 should be the constructor's aim. Applying supply voltage to the top end of R201 provides the 800Hz lf shift necessary on receive. Temperature compensation may be obtained by replacing C201 with a parallel combination of 47pF (polystyrene) and a 22pF npo ceramic.

The basic circuit can be scaled for bands other than 14MHz, or, alternatively, a completely different design utilized. Be careful, however, when selecting a vfo circuit, to ensure that the output amplifier/buffer is capable of delivering 5mW into the 50Ω load presented by the SBL1 mixer.

Transmit/receive control system

In order to attain the operational convenience of a typical commercial transceiver, we must provide some means of enabling the morse key to control automatically the transmit receive changeover function. There are a number of additional facilities that it is desirable to incorporate, and the circuit of Fig 9 is designed to achieve the following:

- (1) The provision of a keying input circuit that will interface with a wide variety of both hand keys and electronic keyers. The standard chosen is a keying potential of 5V positive to ground with the key-down current limited to 100μA.
- (2) A stable changeover timing circuit that provides a variable delay over the range 0.5 to 2s.
- (3) A relay driver to energize the changeover relay coils automatically, as and when required.
- (4) Generation of a keying waveform to feed the transmitter keying transistor.
- (5) Control of the tx/rx offset potential which shifts the vfo frequency 800Hz lf on receive.
- (6) Generation of a sidetone for transmit monitoring.

IC301a functions as a comparator and operates in conjunction with the voltage divider chain comprising R301, R302 and R303. The inverting input of IC301a is held at a positive potential of approximately 5V via R304. The non-inverting input of the op-amp is directly connected to the junction of R302/R303, thus establishing a potential of +2.5V on pin 3. This results in the output of IC301a being forced low, and in practice gives rise to an output of about 1V at pin 1. Depressing the key grounds pin 2 of IC301, with R304 limiting the keying current to 100μA. The non-inverting input of the op-amp is now at a higher potential than the inverting input (ie +2.5V as opposed to 0V) and so the output of IC301a is taken high. The keying waveform is fed to the sidetone oscillator (IC302) via the potential divider R305/R306. The pitch of the sidetone is determined by the values of R307, R308 and C305. Making R307 a preset component therefore provides a means of adjusting the sidetone pitch. The sidetone output may be coupled to the input of the audio output stage via a high-value series resistor, the precise value to be determined by trial and error.

IC301b operates as a monostable and produces the changeover delay in the following manner. Each depression of the key will cause the output of

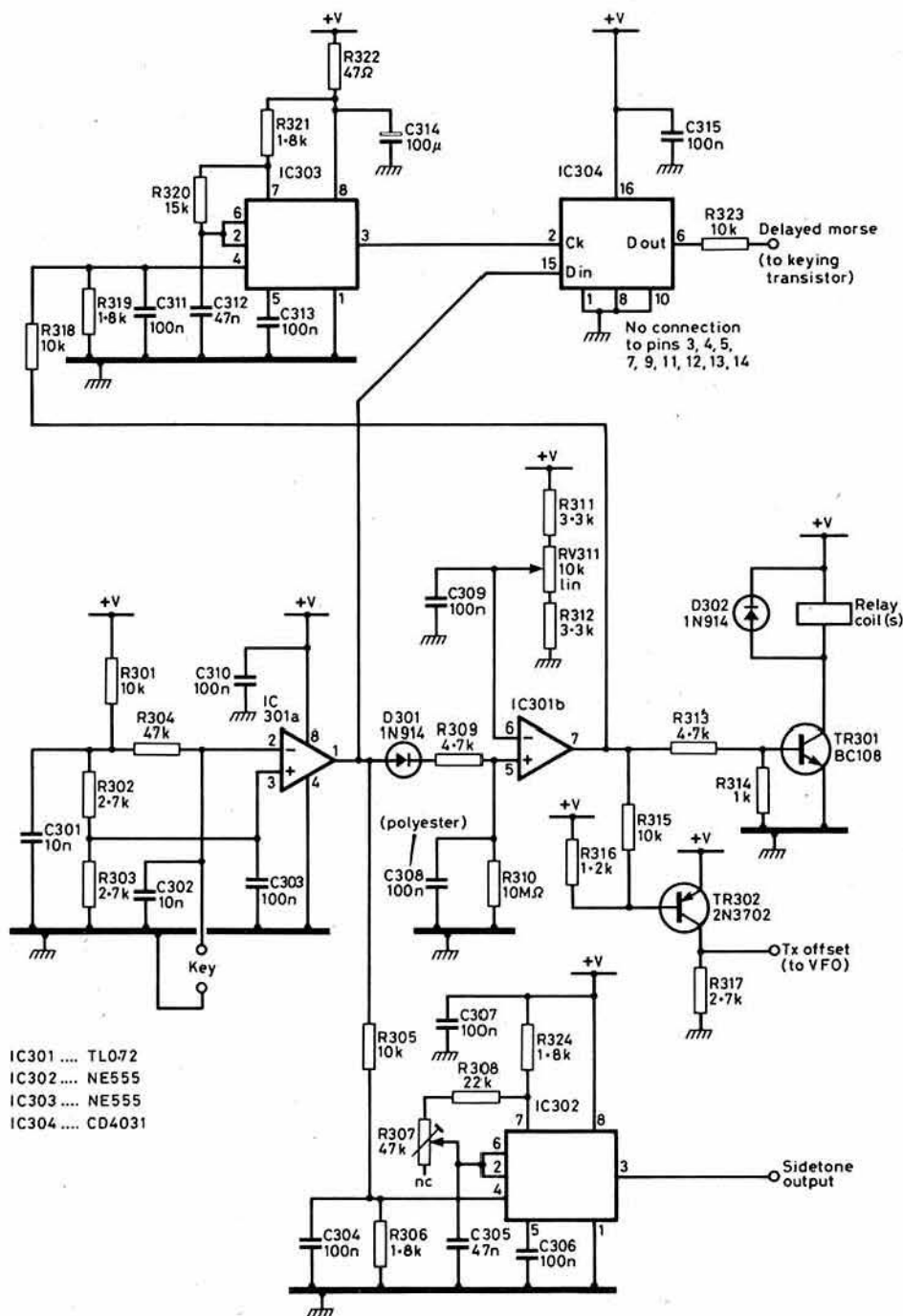


Fig 9. The transmit/receive control system

IC301a to go high, as previously explained. D301 therefore conducts and almost instantaneously charges C308, via R309, to very nearly rail voltage. The positive potential now present at the non-inverting input of IC301b will exceed the potential developed by the variable divider network (R311, R312 and RV301) at the inverting input of the op-amp. This causes the output of IC301b (pin 7) to go high and so turns on the relay driver (TR301) via R313/R314. C308 will stay almost fully charged for as long as morse is being sent, thus ensuring that the changeover relay coils remain energized. When the operator ceases sending, C308 begins slowly to discharge through the high-value resistor R310. Eventually the voltage remaining across C308 will fall below the voltage on pin 6 of IC301b. At this point the op-amp output drops quickly to about 1V, and TR301 turns off, thus de-energizing the relays. RV301 controls the changeover delay time.

TR302 acts as an inverter to create the necessary control voltage for the vfo offset (see previous section). Morse appearing at pin 1 of IC301a is also fed to the data input of a 64-stage static-shift register type CD4031 (IC304). Morse is clocked through the shift register at a predetermined rate and reappears at the output (pin 6) approximately 50ms later. Transmitter keying

Components list

R1	0.22Ω 2.5W w/w	TR1, 3	BC183L
R2, 3	1.8kΩ 0.4W metal film	TR2, 4	2N2905
R4, 6, 8, 9	10kΩ 0.4W metal film	TR5, 6	2N3055
R10	220Ω 0.4W metal film		
R11	1kΩ 0.4W metal film	DI-4	Bridge PWO5
R12	47kΩ 0.4W metal film	D5, 6, 7	IN4148
R13, 14	150kΩ 0.4W metal film	D8, 9	IN4004
R15	270Ω 2.5W w/w	D10	IN4148
RV1	1MΩ preset		
RV2	47kΩ preset	ZD1, 2	BZY88C15
RV3	220kΩ panel mtg pot	IC1	741 op-amp
T1	LT44 (Eagle)	RLA	OM1 12V pcb mtg
C1, 2	1,000μF axial electrolytic	FS1, 2, 3	5A 20mm
C3	1μF 10V axial electrolytic		
C4	0.047μF polyester or similar		
C5	Reversible electrolytic (see text)		

Miscellaneous Eight-way pcb connector, six 20mm pcb fuse clips, pcb, two heatsinks 7.2°C/W, metal box, knob, assorted hardware

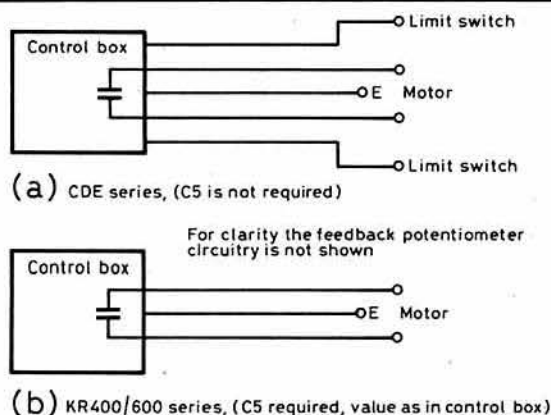


Fig 3. Rotator wiring diagram

Setting up and testing

Consult Fig 4 and wire up the unit accordingly:

(a) CDE series:

- (1) Disconnect the two wires coming from the limit switches and connect these to the output terminals 1 and 2 of the unit.
- (2) Connect two wires from the unit input terminals to where the two wires were removed in (1).
- (3) Connect an earth wire from the control box to the unit.

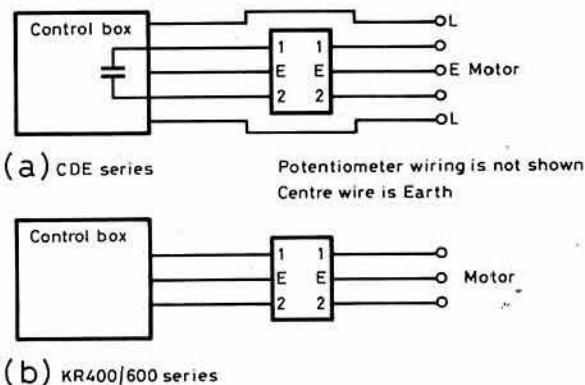


Fig 4. Connection details

(b) KR400/600 series:

- (1) Remove the two wires coming from the motor and connect these to the output terminals of the unit.
- (2) Connect two wires from the unit input terminals to where the two wires were removed in (1).
- (3) Connect an earth wire from the control box to the unit.
Turn RV3 fully clockwise (maximum speed) and adjust RV2 for the maximum speed, consistent with even running of the rotator. Turn RV3 fully anticlockwise and adjust RV1 for the slowest speed, consistent with even running.

The following points should be noted:

- (1) RV1 and RV2 need to be set with the antennas on the rotator.
- (2) The starting torque at higher speed settings will be much less than at normal speed. Therefore, the motor could fail to commence rotation and this might lead to its being burned-out. It is sensible, therefore, to start at low speed and progressively increase to the desired setting. This problem is unlikely to occur except in the case of very large arrays.
- (3) Starting at low speeds and accelerating to that desired will greatly reduce the strain imposed on the rotator. Similarly, decelerating to the required bearing will help prevent overshoot, leading to greater accuracy in the positioning of the antenna system.

In my case, the unit has certainly proved its worth to the extent that I wonder how I ever managed without it!

References

- [1] Details of the kit may be obtained from G3ZUM by sending an s.a.c. to the address given on page 32.
- [2] Maplin Electronic Supplies Ltd, PO Box 3, Rayleigh, Essex.

DIRECT-CONVERSION CW TRANSCEIVERS

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Taking things further . . .

Multiband operation may be achieved by providing a separate, switch-selected, receiver input filter and complementary transmitter lowpass filter for each band. Obtaining vfo drive on more than one band is less straightforward. Nevertheless, with careful design it should be possible to produce a bandswitched vfo of the type employed in the Atlas 210x and 215x transceivers [5]. Other alternatives include the use of a frequency multiplier chain driven by the vfo (for harmonically-related bands), or a simple pre-mix synthesizer using the vfo in conjunction with a crystal oscillator and mixer.

Looking at the project in a wider context leads to the realization that d-c transceivers are not necessarily limited to cw. By extending the vfo

offset facility it is feasible to provide for various fsk/fm modes, including rtty, amtor, sstv and data—perhaps in conjunction with a home computer.

References

- [1] "The G4BWE audio filter". *Rad Com* November 1981, pp1016-9.
- [2] "The G4BWE cw filter". *Rad Com* March 1983, pp226-9.
- [3] "The G4BWE tunable cw filter". *Rad Com* September 1984, pp775-8.
- [4] *Solid State Design For The Radio Amateur*, Wes Hayward, W7ZOI, and Doug De Maw, W1FB. Published by ARRL.
- [5] *Atlas 210X/215X Owners Handbook*.
- [6] *Active Filter Cookbook*, Don Lancaster. Published by H W Sams.

Suppliers

Resistors and capacitors obtainable from Electrovalue Ltd, 28 St Judes Road, Englefield Green, Egham, Surrey TW20 0HB.

Semiconductors obtainable from Technomatic Ltd, 17 Burnley Road, London NW10 1ED.

Amidon toroid cores obtainable from Bonex Ltd, 102 Churchfield Road, Acton, London W3 6DH.